

250-MHz DISTRIBUTED AMPLIFIER SUITABLE FOR CATV TRUNK LINE APPLICATIONS

by Larry Blaser and Norman Sturn

INTRODUCTION

A CATV (Community Antenna Television) system is a means by which a paying subscriber obtains television signals and (often) FM from a centrally located antenna system. Trunk line amplifiers are used every certain cable length to compensate for the normal signal attenuation of the particular type of cable used.

These amplifiers must be capable of handling the broad TV bandpass of 50 to 220 MHz and able to compensate for the bandpass "tilt" caused by increased attenuation of the cable at the higher frequencies. Typical performance requirements^{1,2} are as follows: Gain at 216 MHz should be 25 db or more. The noise figure should be less than 13 db and would be determined by the input stages of the amplifier. The output capability should exceed 45 dbmV (0 dbmV = 1 mV at 75 Ω) and is almost always limited by the cross-modulation introduced by the output stages. The additive nature of the noise figure and cross-modulation limit the number of amplifiers that can be cascaded, and hence, the size of the system.

The design of a wide band amplifier which has a low cross-modulation index and is suitable for the output stage of a trunk line amplifier will be described in this paper. This amplifier would also be suitable for use in the output stage of CATV line extension amplifiers and could be considered for application in other wide-band, high-level, high information-handling systems.

DISTRIBUTED AMPLIFIER

The target design is for an amplifier with a minimum bandwidth of 250 MHz giving a gain of 6 db and output capability of 60 dbmV (1V at 75 Ω) with greater than 60-db cross-modulation rejection.

1. See list of references at the end of this bulletin.

Based on previous experience, it was clear that this performance would not be obtained from a single low-cost bipolar transistor. Instead it was decided that a distributed amplifier design should be investigated. A simple amplifier was considered using inexpensive SE5025 transistors which have an f_t in excess of 500 MHz at I_C equal to 20 milliamperes and V_{CE} equal to 10 volts. These bias conditions are nearly optimum for linear operation of the SE5025 transistor at high output level.

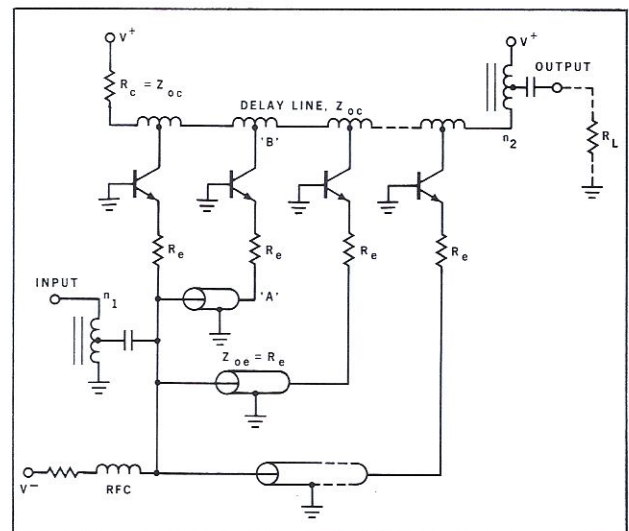


Fig. 1. Basic distributed amplifier configuration.

The basic distributed amplifier stage shown in Figure 1 contains a number of transistor sections with each transistor collector connected in a conventional manner into an artificial delay line. The input is fed through separate

COPYRIGHT, FAIRCHILD SEMICONDUCTOR, 1967 • PRINTED IN U.S.A. 2320-242-27 35M

313 FAIRCHILD DRIVE, MOUNTAIN VIEW, CALIFORNIA, (415) 962-5011, TWX: 910-379-6435

FAIRCHILD
SEMICONDUCTOR
A DIVISION OF FAIRCHILD CAMERA AND INSTRUMENT CORPORATION

Fairchild cannot assume responsibility for use of any circuitry described. No circuit patent licenses are implied.

coaxial delay lines to the emitters of each of the transistors (except the first). Wide-band transformers are used at the input and output for impedance matching.

A simplified expression for the power gain of the amplifier stage is given by:

$$A_p = \left(\frac{g_m \frac{Z_{oc}}{2} N}{n_1 n_2} \right)^2 \frac{R_L}{R_i} \quad (1)$$

where: g_m = forward transconductance between points "A" and "B"

Z_{oc} = Characteristic impedance of collector delay line

N = number of sections

n_1 = turns ratio of input matching transformer

n_2 = turns ratio of output matching transformer

R_L = load resistance

R_i = amplifier input resistance.

The turns ratios, n_1 and n_2 , of the matching transformers are given by:

$$n_1 = \sqrt{\frac{NR_i}{R_e}} \quad (2)$$

$$n_2 = \sqrt{\frac{Z_{oc}}{R_L}} \quad (3)$$

The forward transconductance, g_m is:

$$g_m = \frac{1}{R_e} \quad (4)$$

with the assumption that the common-base input impedance is small compared to the value of R_e .

Substituting the above values of n_1 , n_2 , and g_m into Equation 1 gives:

$$A_p = \frac{N}{4} \frac{Z_{oc}}{R_e} \left(\frac{R_L}{R_i} \right)^2 \quad (5)$$

If R_i is made equal to R_L , then:

$$A_p = \frac{N}{4} \frac{Z_{oc}}{R_e} \quad (6)$$

Equation 6 shows that the maximum power gain is obtained by maximizing N and Z_{oc} and by minimizing R_e . The maximum characteristic impedance of the collector delay line is limited by the required cutoff frequency and by the transistor output capacitance and associated strays. To obtain a constant and linear g_m over the frequency band (i.e., for Equation 4 to be valid), R_e must be large compared to the transistor common-base input impedance. R_e will therefore have some minimum value. The actual values of N , R_c , and R_e must be chosen to be compatible with practical transformer turns ratios.

A four-section amplifier based on the configuration shown in Figure 1 was constructed using transformer turns

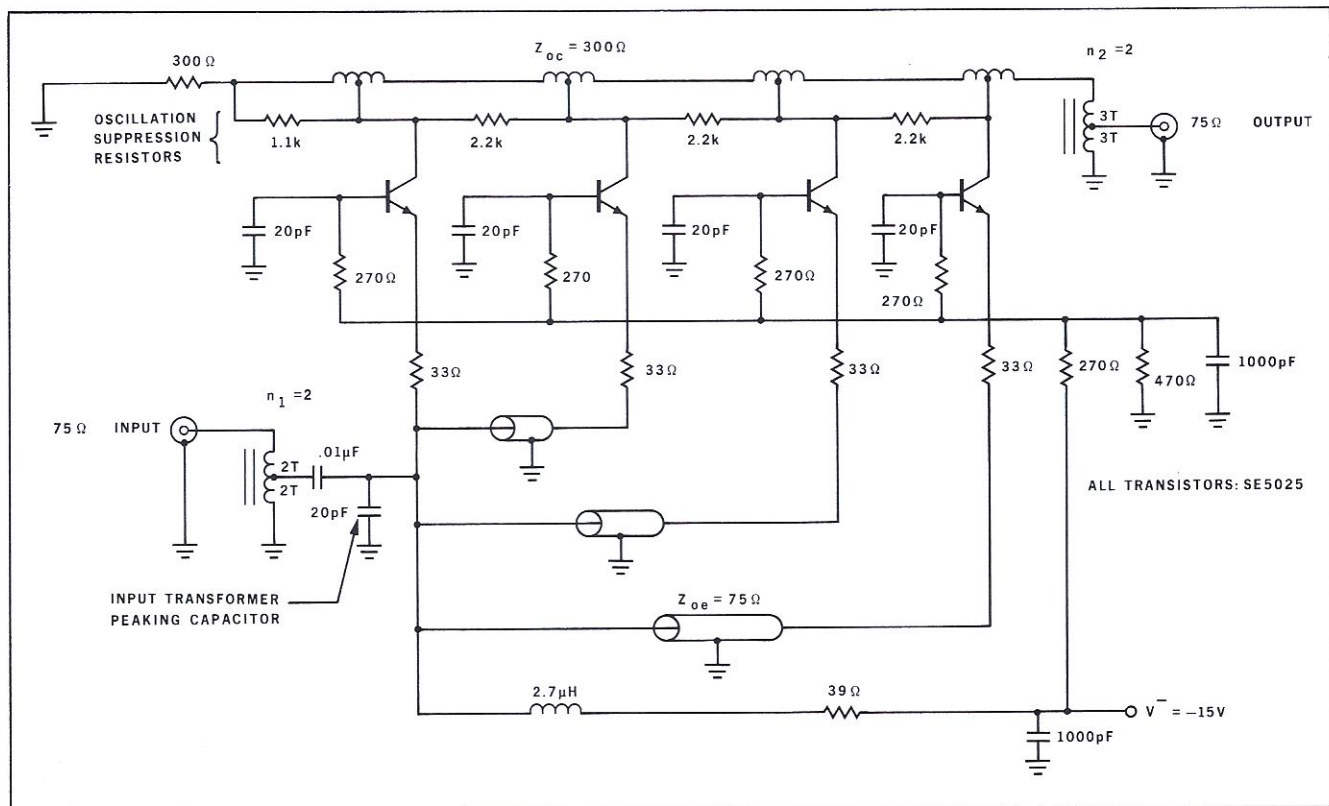


Fig. 2. Distributed amplifier with base compensation.

ratios of 2:1 with input and output impedances of 75 ohms. The collector delay line impedance is 300 ohms and the emitter resistors are each 75 ohms. From Equation 6, the expected gain is $A_p = 4$ (6 db). Experimental results however, showed a fall off in gain at the high end of the TV VHF band. The cause was attributed to the transistor common-base input impedance at these frequencies not being small compared to the 75-ohm emitter resistor.

In a new circuit configuration, shown in Figure 2, a base compensation network was added to extend the response of the amplifier. The values of the base compensation components and of resistor R_e were selected empirically to obtain both a relatively constant g_m between points "A" and "B" and a nearly uniform 75 ohm impedance measured into point "A" for all frequencies in the TV VHF band.

DESIGN DETAILS

The transistor output capacitance, C_{ob} , and the associated stray capacitances are fairly well defined and are the only capacitances used to make up the collector delay line. The sections of the delay line are designed for flattest group delay to best match the constant group delay of the coaxial delay lines that feed the emitters. Flattest group delay is achieved using m-derived sections with m equal to 1.27³. For m equal to 1.27, the delay is essentially flat to 0.6 of the cut off frequency of the line.

Referring to Figure 3, the design of the delay line is as follows:

Given:

$$C_{ob} = 1.8 \text{ pf}$$

$$C_{stray} = 1 \text{ pf}$$

$$m = 1.27$$

$$Z_{oc} = 300\Omega$$

Then:

$$C_1 = C_{ob} + C_{stray} = 2.8 \text{ pf}$$

$$C = \frac{C_1}{m} = 2.2 \text{ pf}$$

$$L = Z_{oc}^2 C = 0.198 \mu\text{h}$$

$$L_1 = \frac{m^2 + 1}{4m} L = 0.103 \mu\text{h}$$

$$M = \frac{m^2 - 1}{4m} L = 0.02 \mu\text{h}$$

$$f_c = \frac{1}{\pi \sqrt{LC}} = 480 \text{ MHz (delay line cut off frequency)}$$

$$0.6 f_c = 290 \text{ MHz (frequency to which group delay is flat)}$$

$$t_s = m \sqrt{LC} = 0.83 \text{ nsec (time delay per section)}$$

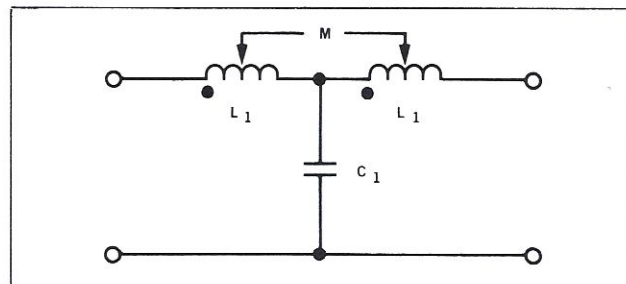


Fig. 3. An m-derived section of delay line.

An RG 187u, 75-ohm coaxial cable is used in the emitters to match the delay in the collector line. The measured delay of the RG 187u cable is 0.133 nanoseconds per inch. The shortest cable (0.83 nsec) is therefore cut to 6.25 inches. The other two cables are twice and three times this length.

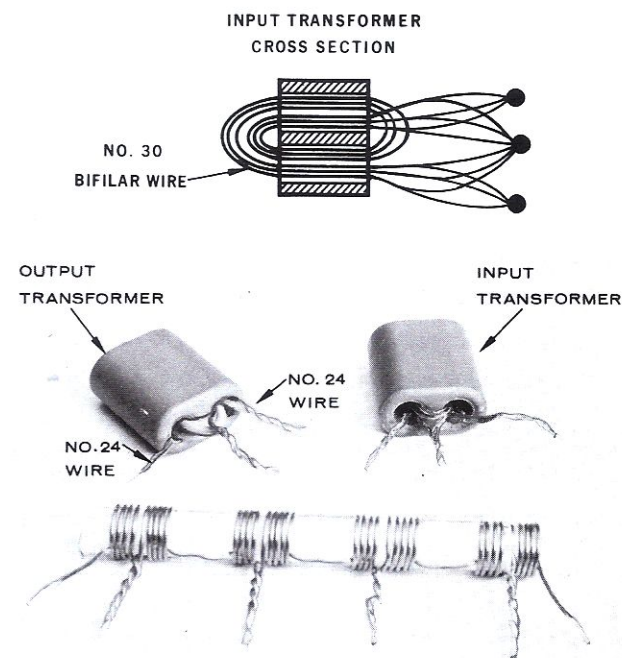


Fig. 4. Transformers and delay line.

The windings of the 4:1 impedance transformers⁴ at the input and output of the amplifier are, in themselves,

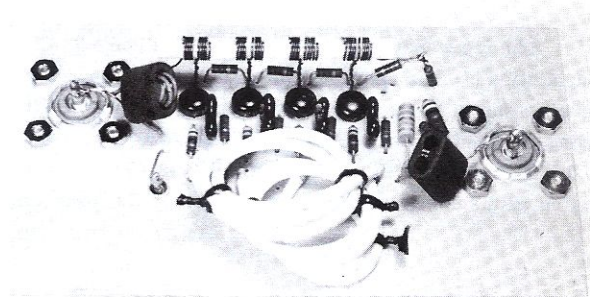


Fig. 5. Distributed amplifier stage.

transmissions lines which are very broadband. The characteristic impedance of the wire constituting the windings is equal to the secondary loading.

A photograph of the collector delay line and of the input and output transformers is shown in Figure 4. The complete amplifier is shown in Figure 5. A detailed drawing of the collector delay line is shown in Figure 6.

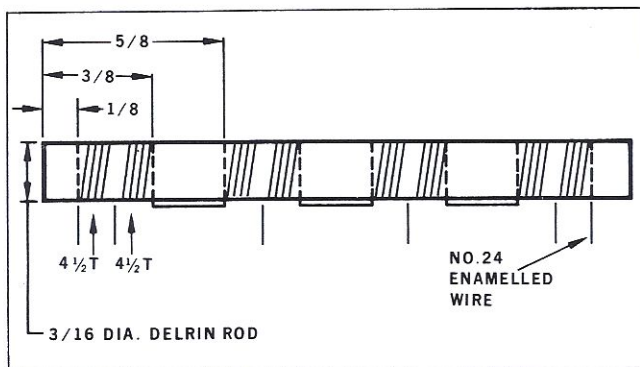


Fig. 6. Delay line.

PERFORMANCE

The performance characteristics for one section of the distributed amplifier are listed below:

- Transducer Gain = 6 ± 0.5 db, 50 to 200 MHz
- Input Impedance = 75 ± 25 ohm, 50 to 220 MHz
- Output Impedance = 75 ± 25 ohm, 50 to 220 MHz
- Power Supply = -15V, $I_{SS} = 100$ mA.

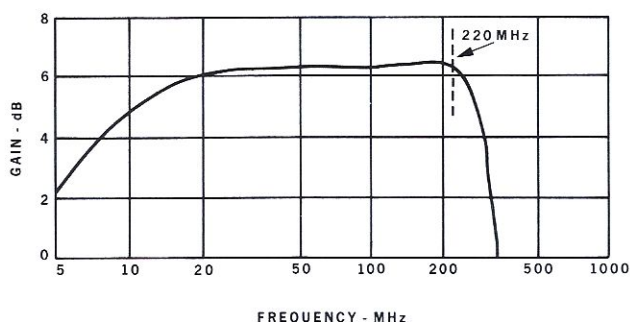


Fig. 7. Amplifier frequency response.

The frequency response is shown in Figure 7. Cross-modulation of four cascaded amplifier stages versus output level is shown in Figure 8. These measurements were made using a two-frequency system. The two carriers were 160 MHz and 200 MHz, each of which was 15 percent modulated with a 1-kHz square wave. Each of the four ampli-

fier stages was operated at the same level by placing 6-db pads between the stages. Cross modulation of a single amplifier, except at very high output levels, could not be measured accurately because of nonlinearities in the measurement system. Even with four cascaded amplifiers it was not possible to ascertain the cross-modulation rejection at outputs below 50 dbmv. To obtain values of cross-modulation rejection below 50 dbmv, it would be necessary to cascade more amplifiers and/or improve the cross-modulation rejection of the measurement system.

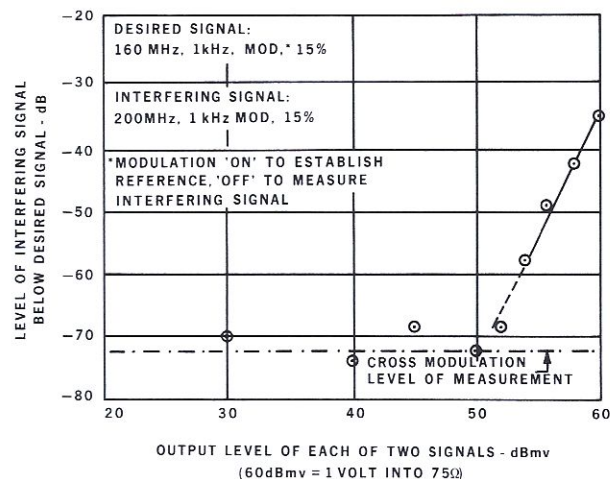


Fig. 8. Cross modulation rejection of four cascaded amplifiers, each operated at the same output levels.

CONCLUSION

An amplifier has been described which has a greater output capability than the final stage of conventional CATV trunk line amplifiers. Use of the described amplifier would improve overall CATV systems performance and allow more subscribers to be reached. Fewer amplifiers would be required per mile of cable.

REFERENCES

1. R. D. Chipp, "Community Antenna Television Systems", *IEEE Spectrum*; July 1966.
2. M. G. Richey, "A Review of Community Antenna Television Trunk Line Design Considerations", presented at the 1966 IEEE International Communications Conference, Philadelphia, Pa. June 15-17, 1966 Paper No. CP-19CP66-1120.
3. Millman and Taub, "Pulse and Digital Circuits" McGraw-Hill Book Company, Inc. pp 291-298; 1956.
4. C. L. Ruthroff, "Some Broad-Band Transformers", *Proceedings of the IEEE*; August 1959.